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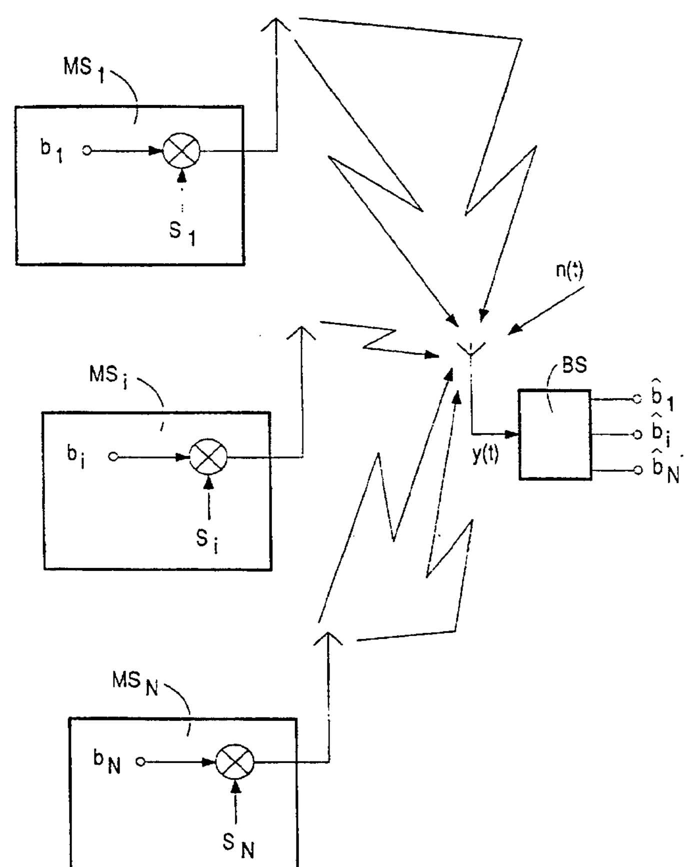
(54) Title: CDMA COMMUNICATION SYSTEM

(57) Abstract

The invention relates to a CDMA communication system comprising at least a transmitter in which a data sequence is spread with the code sequence and comprising at least a receiver in which the data sequences are recovered by a detector, estimate sequences for the values of the transmitted data sequences being computed for the purpose of the detection while a description of the communication channel between the transmitter or transmitters and a receiver is incorporated. For the detection of data sequences there is provided to compute joint probability distributions for the transmitted data sequences by computing the associated moments and determine therefrom the estimate sequences for the transmitted data sequences.



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CDMA communication system.

The invention relates to a CDMA communication system comprising at least a transmitter in which a data sequence is spread by a code sequence and comprising at least a receiver in which the data sequences are recovered by a detector, sequences of estimates for the transmitted data sequences being computed for the purpose of the detection while a description of the communication channel situated between the transmitter or transmitters and a receiver is incorporated. The invention likewise relates to a receiver for a CDMA communication system and, more specifically, to a detector for such a receiver.

In a communication system based on Code-Division Multiple Access (CDMA) the signals of various users are simultaneously transmitted in a common frequency band and with a common carrier frequency. Code-division multiple access systems are based on a spread band technique, that is to say, the signal to be transmitted is spread out over a frequency band which is substantially wider than the minimum frequency band necessary for transmitting the signal. As a result of the spreading of the band, code-division multiple access systems are generally highly insensitive to interference.

For frequency band spreading in digital communication systems, for example, each bit to be transmitted is multiplied by a codeword mutually agreed upon by transmitter and receiver. When mutually orthogonal codewords are used, mutual interference by the signals from the individual users is ruled out in principle. But realistic requirements as to wave propagation over the surface of the earth such as multipath propagation lead to the fact that this orthogonality is no longer adhered to. If, in addition, a synchronization of the accesses by the individual users is dispensed with to realise a highly simple access to the common frequency band, or if different bit rates are permitted, the signal of a user can be detected only with more circuitry and cost or with a degraded quality, because the signals can no longer be mutually orthogonal.

From "Linear Multiuser Detectors for Synchronous Code-Division Multiple-Access Channels" by Ruxandra Lupas and Sergio Verdú, published in IEEE

Transactions on Information Theory, vol. 35, no. 1, January 1989, pp. 123-136, are already known detectors for a CDMA receiver in which detectors various users are detected simultaneously. By simultaneously detecting various users, the quality of the detection can be improved by increasing the circuitry in that the cross-correlation between the signal portions of different users is eliminated or reduced by a linear image. In consequence, also interference caused by receiver noise is amplified. For the rest such detectors are sub-optimum as regards bit error rate.

"Minimum Probability of Error for Asynchronous Gaussian Multiple-Access Channels" by Sergio Verdú, published in IEEE Transactions on Information Theory, vol. IT-32, no. 1, January 1986, pp. 85-96, describes non-linear detectors which are nearly optimal with respect to the bit error rate. But the cost of realising such ideal non-linear detectors rises exponentially with the number of users of the CDMA communication system.

It is an object of the invention to provide a detector for CDMA communication systems whose performance relative to linear detectors is improved, but whose cost of realization remains below the cost of an optimum detector.

This object is achieved in that joint probability distributions for the transmitted data sequences are computed by computing the associated moments and deriving therefrom the sequences of estimates for the transmitted data sequences.

During communication, new data are continuously transmitted and the receiver makes decisions about the transmitted data values. In consequence, the joint probability distribution changes continuously, so that a sequence of joint probability distributions is to be computed. In addition, the joint probability distribution can be constantly improved by the incoming receive signal. When joint probability distributions are computed, the information about the communication channel and the properties of noise can be fully taken into account.

Furthermore, probability distributions automatically provide quality information about the estimates. This quality information can be evaluated beneficially in a decoder downstream in the circuit.

The use of moments to describe probability distributions is advantageous in that a highly simple description of the probability distribution can be obtained with the moments.

In a further embodiment of the invention the joint probability distribution

is computed only with a restricted set of moments.

A restricted set of moments will provide only an approximate description of the joint probability distributions. The use of a restricted set of moments, however, provides a reduction of cost of the detector, without having to accept significant quality losses on detection.

In a further embodiment of the invention only first and second order moments of the joint probability distribution are computed to compute an approximate joint probability distribution.

It has appeared that, as a rule, it is sufficient for a substantially optimum detection to compute the first and second moments of the joint probability distribution, thus to compute the mean value and the covariance.

In a special embodiment of the invention a Kalman filter is provided as a detector.

The use of a modified Kalman filter for what is commonly referred to as a soft decision feedback equalizer has been examined in "A NEW NONLINEAR EQUALIZER FOR MOBILE RADIO CHANNELS", J. Tielecke, Proceedings ICASSP 90 (INTERNATIONAL CONFERENCE ON ACOUSTICS, SPEECH AND SIGNAL PROCESSING; APRIL 3-6, 1990) pp. 1667-1670, for conventional TDMA communication systems to eliminate intersymbol interference. In TDMA systems the time-dispersive communication channels predominantly cause intersymbol interference to occur. (In the case of intersymbol interference, mutually time shifted bits of the same user are superimposed on each other.) In CDMA systems this intersymbol interference plays only a minor role compared to the interference caused by other users. For the man skilled in the art said examination did not produce any indication to apply this Kalman filter technique also to CDMA systems and their different types of problems.

Customarily, with a Kalman filter the computed estimate vector and associated error covariance matrix are not assigned to a probability distribution. However, as is known, the estimate vector can be considered a mean vector of the joint probability distribution (first moment of the joint probability distribution) and the error covariance matrix can be considered a covariance matrix of the joint probability distribution (second moment of the joint probability distribution.

In a further embodiment of the invention decisions about the values of the data sequence are made on the basis of the estimates and these decisions are fed back

inside the detector.

As a result of the decisions that have been fed back the quality of the detection process can generally be improved drastically, because the discrete nature of the transmitted data sequences is taken into account. Especially with a Kalman filter such a feedback does not take place.

In a further embodiment of the invention the quality of the estimate, when the values are fed back, is taken into account for the value that has been fed back.

By avoiding hard decisions, the quality of the detection process can be improved even more. If the detector is followed by a decoder in the circuit, which decoder can process decisions with quality grading (soft decision), the communication reliability is enhanced distinctly.

The invention will now be further described and explained with reference to the embodiments represented in the drawing Figures, in which:

Fig. 1 shows a CDMA communication system comprising N mobile stations and one base station,

Fig. 2 shows a receiver for such a CDMA communication system,

Fig. 3 shows function blocks of a detector,

Fig. 4 shows a time diagram representing transmitted data sequences of a plurality of users of a CDMA communication system,

Fig. 5 shows a state diagram of a modified Kalman filter for estimating transmitted data signals,

Fig. 6 shows a block circuit diagram of a feedback step in the state diagram shown in Fig. 5,

Fig. 7 shows a state diagram for the estimation of the covariance matrix, and

Fig. 8 shows a block circuit diagram of a feedback step in the state diagram shown in Fig. 7.

In the CDMA data communication system shown in Fig. 1 the transmit data are available as binary data sequences b_i. In the embodiment the binary data sequences b_i are recovered from speech signals by appropriate coding. These binary data sequences b_i are exchanged between individual mobile radio sets and their base station BS in whose coverage area the mobile radio sets happen to be located. N data sources b_i are available which corresponds with the number of N simultaneously

transmitting mobile stations, while the subscript index i = 1 ... N in the drawing Figures forms the distinction between the individual data sources. To spread a data sequence, each data sequence b_i is multiplied by a code sequence s_i. In the following a bit of the code sequence is referenced a chip to distinguish it from a bit of the transmitted data sequence.

For the transmitted signals to be separated on the receiving side, the individual code sequences are to be selected discriminatively. For this purpose, for example the mobile station randomly selects a code sequence and transmits the selection to the base station when a connection is established between mobile station and base station.

The different locations of the individual mobile stations provides that the transmitted signals arrive at the base station by different radio channels. By these radio channels the transmitted signals are exposed to distortions caused by, for example, reflection and multipath propagation. These distorted signals are superimposed on each other at the antenna of the base station BS to form a continuous receive signal y(t) which receive signal contains noise signal portions n(t).

In the receiver of the base station BS a receive data sequence $\hat{b}_i$ is estimated for each transmitted data sequence b_i in the received signal y(t). For example, the recovery of speech signals from these input data and the distribution of the received data to the individual receive ports, for example, by transfer to a communication network, are not further shown here.

Fig. 2 shows the basic structure of a receiver according to the invention. The received signal y(t) is first preamplified by a HF preamplifier stage 21 and bandpass filtered. By means of a HF signal coming from a HF oscillator 22, which HF signal corresponds to the carrier frequency used in the transmitter, the bandpass filtered received signal is mixed in mixers 23a, 23b with the HF signal itself and with a signal that has a quadrature phase relative to the HF signal. As a result, the real input signal y(t) is directly converted to the baseband during which operation a complex baseband signal containing a real portion and an imaginary portion is produced. These two signals are sampled at equidistant instants kT by samplers 24a, 24b once they have been low-pass filtered, which filtering is not shown in Fig. 2. The sampling theorem is then to be considered *i.e.* the sampling frequency is to be twice as high as the limit frequency of the baseband signal. In the embodiment this is achieved in that the sampling is carried

out at twice the chip rate.

The samples are subsequently converted in analog-digital converters 25a, 25b into a digital value sequence y(k). This digital value sequence y(k) is applied to a digital signal processor 26 which contains, for example in a read-only memory 27, a program for estimating the transmitted binary data sequences. Intermediate results arising from the estimation are buffered by the signal processor 26 in a read write memory 28. Depending on the number of users to be detected, the computational capacity of a single signal processor is no longer sufficient under specific circumstances. In that case more signal processors over which the computational capacity is to be spread are to be provided.

Fig. 3 shows in a circuit diagram the functions realised by this or by these signal processor(s) respectively, as long as they are necessary for the signal estimation. To perform the signal estimation in the detector 30 a description of the communication channel is necessary for each user, which description is produced by a channel estimator 31. There should be observed in this respect that, due to the different locations of the mobile stations, each radio channel between mobile station and base station is different. A so-termed channel impulse response is often used for the channel description. To determine the channel description it is possible, for example, to insert a training data sequence into the transmit signal, from which sequence the channel impulse response can be computed in the receiver by correlators. Furthermore, a code sequence generator 32 is necessary which produces the value of one chip for each individual user at the sampling instant k. As will further be shown, a channel impulse response formed by a radio channel impulse response and the code sequences will be advantageously made available to the detector 30.

For computing the sequence of moments of the joint probability distributions it is advantageous to use a description of the communication channel in the state space. For this purpose, the data of the individual users which have an effect on the discrete time received signal y(k) at the sampling instant k are combined in a vector $\underline{b}(k)$. As not only bits from different users can cause mutual interference in time dispersive communication channels, but also successive bits from one user, the vector $\underline{b}(k)$ comprises not only one bit per user but, depending on the time dispersion, various bits per user. Assuming the simplification of a linear communication channel which is distorted by additive noise, the following observation equation can be used for the input

signal y(k) observed in the receiver at the sampling instant k:

$$y(k) = \underline{h}^T(k)\ \underline{b}(k) + n(k)$$

The vector $\underline{h}(k)$ describes how the transmitted bits cause interference to one another. The vector takes the actual radio channel, filter in the communication circuit and the spread sequences into account.

To clarify the influence of the code sequences on $\underline{h}(k)$, Fig. 4 shows a very simple embodiment. In this embodiment the CDMA communication system has three users. The modulation method is Phase Shift Keying and the radio channel for each user comprises a direct link (no multipath propagation, no fading, no attenuation), which link is only disturbed by Additive White Gaussian Noise (AWGN). For the instant k_0 the following is obtained for the vector $\underline{b}(k_0)$ which vector comprises the transmitted bits which cause interference to one another

$$\underline{b}(k_0) = [+1\ ,\ +1\ ,\ -1]^T.$$

The impulse response vector $\underline{h}(k_0)$ is in this simple case determined only by the code sequences used for the spreading. From Fig. 4 it follows that:

$$\underline{h}(k_0) = [+1\ ,\ -1\ ,\ +1]^T.$$

As the values of the chips change continuously, a new impulse response vector $\underline{h}(k)$ is to be computed for practically every new sampling instant, which vector takes into account the current values of the chips.

The incorporation of the code sequences in the impulse response vector $\underline{h}(k)$ is advantageous because, as a result, the detector may have a highly flexible structure: For example, code sequences can be permitted of which the period does not correspond to the data period used. Furthermore, the users can be permitted to have different and variable data rates. The embodiment shown in Fig. 4 shows that integrating binary code sequences into the impulse response vector $\underline{h}(k)$ is very simple, because in the impulse vector $\underline{h}(k)$ the signs may only be changed depending on the chips.

The observation equation can be completed by a state transition equation which describes how the composition of the vector $\underline{b}(k)$ is changed at the transition from sampling instant k to sampling instant k+1

$$\underline{b}(k+1) = \underline{A}(k) * \underline{b}(k) + \underline{b}_{\Delta}(k+1)$$

With the aid of matrix $\underline{A}(k)$ there is formally described which data are omitted from the vector $\underline{b}(k)$ or $\underline{b}(k+1)$ as they no longer contribute to the received signal $y(k+1)$ at the instant $k+1$. By means of the vector $\underline{b}_{\Delta}(k+1)$ the data are added to the vector $\underline{b}(k)$ or $\underline{b}(k+1)$ respectively, which data have just been transmitted at the instant $k+1$ and thus influence the received signal for the first time.

The state equation in this form is only used for describing in general by way of a formula how data are added to the vector $\underline{b}(k+1)$ or read from this vector. When $\underline{b}(k+1)$ is computed in a signal processor, such operations are preferably not carried out by matrix multiplications and matrix additions, but by specific storing operations. For the case where no new data are transmitted at instant $k+1$, the vector $\underline{b}(k)$ is changed to $\underline{b}(k+1)$ at the transition.

By means of the magnitudes and descriptions introduced by the description of the state space there will now be described which computing operations are to be carried out in the detector to estimate the data. For simplicity there is assumed for the preferred embodiment that binary values are transmitted as data and that the baseband received signal is a real-value signal. In the case at hand of a complex value baseband signal, the real and imaginary portions of a complex sample value $y(k)$ in the preferred embodiment are treated as two separate real-value signals received consecutively. In Kalman filters even slight advantages as to quality are obtained without the circuitry and cost being enhanced.

In the preferred embodiment first and second-order moments are recursively computed *i.e.* mean value vectors $\hat{\underline{b}}(k)$ and $\hat{\underline{b}}_{+}(k)$ as well as associated covariance matrices $\underline{P}(k)$ and $\underline{P}_{+}(k)$. These moments can be assigned to joint probability distributions, which distributions approach the real probability distributions. Fig. 5 shows a section of a state diagram by which the computation steps to be carried out by the signal processor 26 are shown in diagrammatic form. The detector itself may be considered a modified Kalman filter, modified by a feedback of decisions having information about quality (soft decision). The computation of the covariance matrix necessary for estimating the data sequences is given in diagrammatic form in the state diagram shown in Fig. 7. The computations which are to be carried out by the signal processor 26 for the feedback are shown in Fig. 6 and Fig. 8 in the respective state diagrams. For a better comprehension there may be observed that the block featuring a feedback step in Figs. 5 and 6 is designated 10 and is designated 11 in Figs. 7 and 8.

But also without feedback decisions the Kalman filter used according to the invention presents significantly improved properties compared to prior-art sub-optimum detectors, especially with respect to flexibility.

At the start of a data transmission normally training data are transmitted when the connection is established. Therefore, an initialization of the detector for an additional user is simple, because only the appropriate data are to be entered in the mean value vector. In the covariance matrix the associated covariance values are to be set to zero. In this way at instant k the mean value vector $\hat{\underline{b}}_+(k)$ together with the associated covariance matrix $\underline{P}_+(k)$ is known. The signal value y(k) received at instant k has then not yet been evaluated.

An improved estimate or mean value vector $\hat{\underline{b}}(k)$ for the state vector $\underline{b}(k)$ is then computed in that the previous estimate vector $\hat{\underline{b}}_+(k)$ is corrected on the basis of the received value y(k). This is carried out by the filter equation of a Kalman filter, which equation is based on the observation equation of the description of the state space. The following holds for the corrected mean value vector

$$\hat{\underline{b}}(k) = \hat{\underline{b}}_+(k) + \underline{g}(k)\, e(k)$$

The correction term can be computed from the Kalman gain vector

$$\underline{g}(k) = \frac{\underline{P}_+(k)\, \underline{h}(k)}{\underline{h}^T(k)\, \underline{P}_+(k)\, \underline{h}(k) + \sigma_n^2}$$

and from the estimated error e(k) for the received signal y(k)

$$e(k) = y(k) - \underline{h}^T(k)\, \hat{\underline{b}}_+(k) .$$

In the Kalman gain vector g(k) the power σ_n^2 of the receiving noise n(k) is to be taken into account. The power can easily be estimated in the receiver, for example, within the scope of the channel identification. An accurate estimate is not necessary because Kalman filters are, as is known, robust to errors of the noise power.

Just as the mean value vector was corrected, so must the associated covariance matrix be corrected. The corresponding Kalman filter equation reads:

$$\underline{P}(k) = \underline{P}_+(k) - g(k)\,\underline{h}^T(k)\,\underline{P}_+(k)\ ,$$

while

again the Kalman gain vector g(k) was used. The moments $\hat{\underline{b}}(k)$ and $\underline{P}(k)$ are first and second-order moments at which the currently received sample value y(k) is taken into account.

Before changing from sampling instant k to k+1, said feedback (soft decision) is introduced, provided that a bit is to be decided on. A decision is then made when the bit at instant k+1 no longer occurs in the state vector $\underline{b}(k+1)$ of the channel model. Due to the feedback the estimates are improved, because the Kalman filter (without feedback) does not take into account that binary signals are transmitted. The Kalman filter implicitly rather assumes Gaussian probability distributions *i.e.* Gaussian distributed transmitted data. With the feedback there is included in the (Gaussian) probability distribution belonging to $\hat{\underline{b}}(k)$ and $\underline{P}(k)$ that the i^{th} element of the state vector $\underline{b}(k)$ omitted at the transition from k to k+1 is binary. The resulting probability distribution has the moments $\hat{\underline{b}}_*(k)$ and $\underline{P}_*(k)$. It is especially advantageous that in the embodiment the improved moments are computed by equations which are similar to the filter equations of the Kalman filter. As a result, essentially the same algorithm can be used for the computation.

For the mean value vector $\hat{\underline{b}}_*(k)$ is obtained:

$$\hat{\underline{b}}_*(k) = \hat{\underline{b}}(k) + g_*(k)\,e_*(k)$$

with the modified Kalman gain vector

$$g_*(k) = \frac{\underline{P}(k)\,\underline{u}_i}{\underline{u}_i^T\,\underline{P}(k)\,\underline{u}_i}$$

and with the estimate error term $\underline{e}_*(k)$.

$$e_*(k) = \alpha(k) - \underline{u}_i^T \, \hat{\underline{b}}(k) \,.$$

For $\alpha(k)$ may then be taken:

$$\alpha(k) = \tanh \frac{\underline{u}_i^T \, \hat{\underline{b}}(k)}{\underline{u}_i^T \, \underline{P}(k) \, \underline{u}_i} \,.$$

The vector $\underline{u}_i$ is a unit vector in which only the i^{th} element is different from zero *i.e.* is equal to one. It was assumed that the i^{th} element of the state vector $\underline{b}(k)$ is to be decided on and sampled from the vector.

Another embodiment having a "hard" decision about the transmitted data is obtained in that in the equation for $\alpha(k)$ the hyperbolical tangent (tanh) is replaced by the sign function (sign).

For the covariance matrix there is the modified Kalman filter equation

$$\underline{P}_*(k) = \underline{P}(k) - \underline{g}_*(k) \, \underline{u}_i^T \, \underline{P}(k) \, \beta(k)$$

with the correction factor

$$\beta(k) = 1 - \frac{1 - \alpha^2(k)}{\underline{u}_i^T \, \underline{P}(k) \, \underline{u}_i} \,.$$

Thus moments $\hat{\underline{b}}_*(k)$ and $\underline{P}_*(k)$ occur, which take into account that the i^{th} element of the state vector $\underline{b}(k)$ is a bit. Before changing to the sampling instant k+1, and the bit is omitted from the state vector $\underline{b}(k)$, the bit decision is made on the basis of the associated element of the estimate vector or mean value vector $\hat{\underline{b}}_*(k)$. Furthermore, the bit error probability can be approximated because the moments $\hat{\underline{b}}_*(k)$ and $\underline{P}_*(k)$ are assigned to a probability distribution. For the bit error probability there is obtained:

This quality

$$P(\text{bit errors}) = \frac{1 - |\underline{u}_i^T \hat{\underline{b}}_*(k)|}{2}$$

criterion can be advantageously evaluated in a decoder downstream in the circuit, to make the information transmission more reliable.

If a plurality of bits are omitted at the transition from k to k+1, the modified filter equations can be evaluated accordingly often, and instead of $\hat{\underline{b}}(k)$ and $\underline{P}(k)$, the moments $\hat{\underline{b}}_*(k)$ and $\underline{P}_*(k)$ already improved can be inserted. If no bit is omitted, the modified filter equations are cancelled and $\hat{\underline{b}}_*(k) = \hat{\underline{b}}(k)$ and $\underline{P}_*(k) = \underline{P}(k)$ can be stated. This will also be the case if a feedback would generally be dispensed with.

The transition from sampling instant k to k+1 takes place in accordance with the transition equation of the description of the state space. The associated Kalman prediction equation for the mean value vector reads:

$$\hat{\underline{b}}_+(k+1) = \underline{A}(k)\, \hat{\underline{b}}_*(k)\ .$$

For the covariance matrix there is obtained:

$$\underline{P}_+(k+1) = \underline{A}(k)\, \underline{P}_*(k)\, \underline{A}^T(k) + \underline{Q}_{b\Delta}(k)\ .$$

The matrix $\underline{Q}_{b\Delta}(k)$ is a diagonal matrix which contains the variances of the bits that have just been transmitted at instant k+1 *i.e.* a 1 value (otherwise 0) in the row in which the associated bit appears in the unknown vector $\underline{b}(k+1)$. In the corresponding row of the estimate vector or mean value vector $\hat{\underline{b}}_*(k+1)$ there appears a zero because there is assumed that the bits are uniformly distributed *i.e.* without a mean value. With this the prediction step associated with the state transition equation is described and the cycle is closed. In the embodiment no matrix multiplications are carried out in the signal processor during the prediction step, but only storing operations which have already been explained with reference to the state transition equation.

Summarizing there may be stated that the computation cost of the detector

described was drastically reduced compared with the optimum detector, without the bit error rate being increased appreciably. Furthermore, the detector produces estimates for the bit error probabilities. This may be used in a decoder to enhance the reliability of transmission. The detector also uses multipath propagation to advantage. For the code sequences no limitations with respect to the period durations used have to be taken into consideration. Different and variable data rates of the users are permitted. The users need not be synchronized either. All these are advantages which are generally absent in sub-optimum detectors.

Claims:

1. CDMA communication system comprising at least a transmitter in which a data sequence is spread by a code sequence and comprising at least a receiver in which the data sequences are recovered by a detector, sequences of estimates for the transmitted data sequences being computed for the purpose of the detection while a description of the communication channel situated between the transmitter or transmitters and a receiver is incorporated, characterized in that joint probability distributions for the transmitted data sequences are computed by computing the associated moments and deriving therefrom the sequences of estimates for the transmitted data sequences.

2. CDMA system as claimed in Claim 1, characterized in that the joint probability distribution is computed only with a restricted set of moments.

3. CDMA system as claimed in Claim 2, characterized in that only first and second order moments are computed when the joint probability distribution is computed.

4. CDMA system as claimed in Claim 3, characterized in that a Kalman filter is used as a detector.

5. CDMA system as claimed in one of the Claims 1, 2, 3 or 4, characterized in that decisions about the values of the data sequence are made on the basis of the estimates and these decisions are fed back inside the detector.

6. CDMA system as claimed in Claim 5, characterized in that the quality of the estimate, when the values are fed back, is taken into account for the value that has been fed back.

7. Receiver for a CDMA communication system in which receiver data sequences spread by code sequences are recovered by a detector, sequences of estimates for the values of the transmitted data sequences being computed for the purpose of the detection while a description of the communication channel situated between a transmitter or various transmitters and the receiver is incorporated, characterized in that joint probability distributions for the transmitted data sequences are computed by

computing the associated moments and deriving therefrom the sequences of estimates for the transmitted data sequences.

8. Detector for a receiver of a CDMA communication system in which there is provided that the detector computes sequences of estimates for the values of the transmitted data sequences on the basis of a description of the communication channel situated between a transmitter or various transmitters and the receiver, characterized in that joint probability distributions for the transmitted data sequences are computed by computing the associated moments and deriving therefrom the sequences of estimates for the transmitted data sequences.

9. Detector as claimed in Claim 8, characterized in that the code sequences used for the spreading are incorporated in the channel description.

10. Detector as claimed in Claim 9, characterized in that for incorporating the code sequences in the channel description there is provided to carry out sign operations for the channel description.

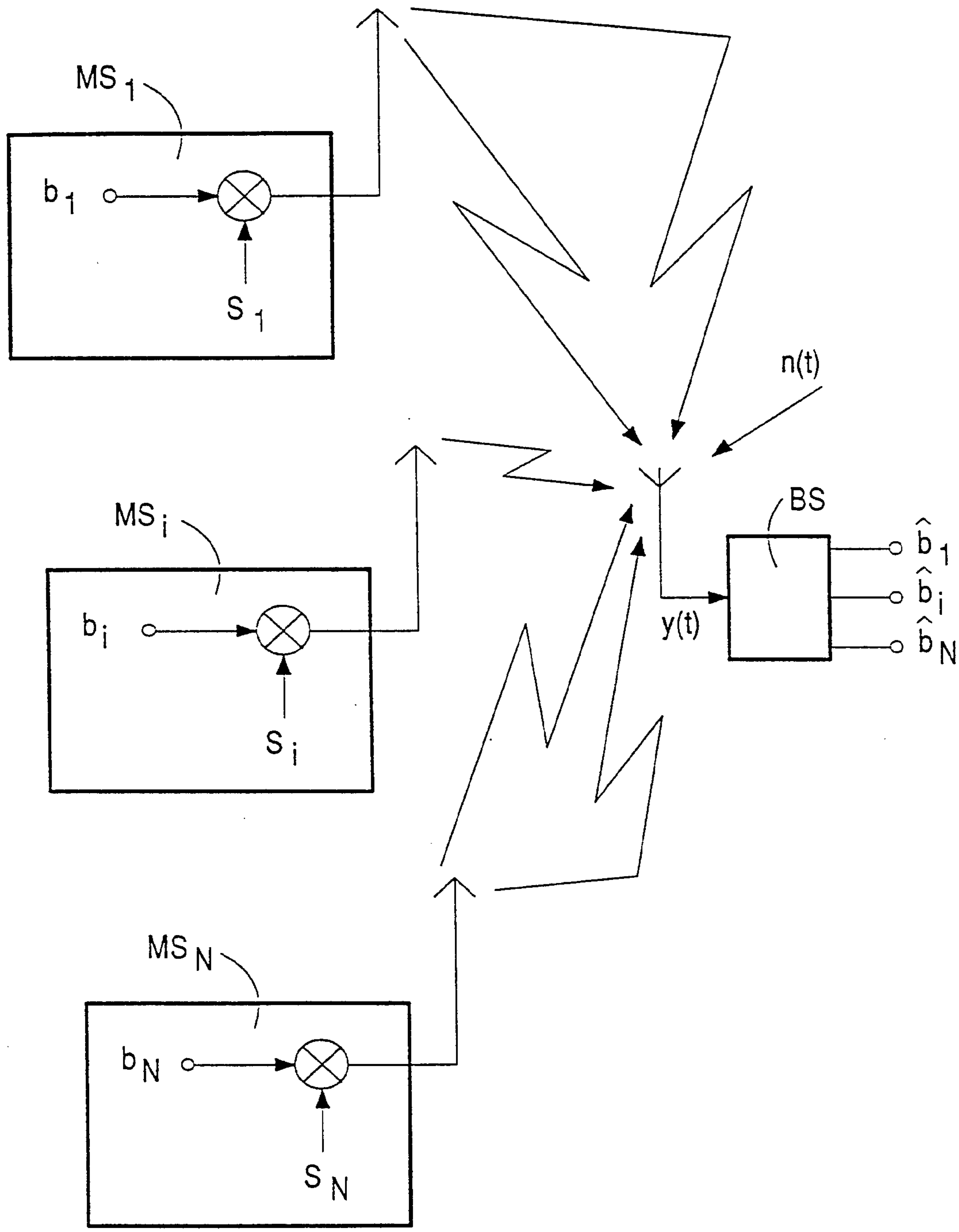
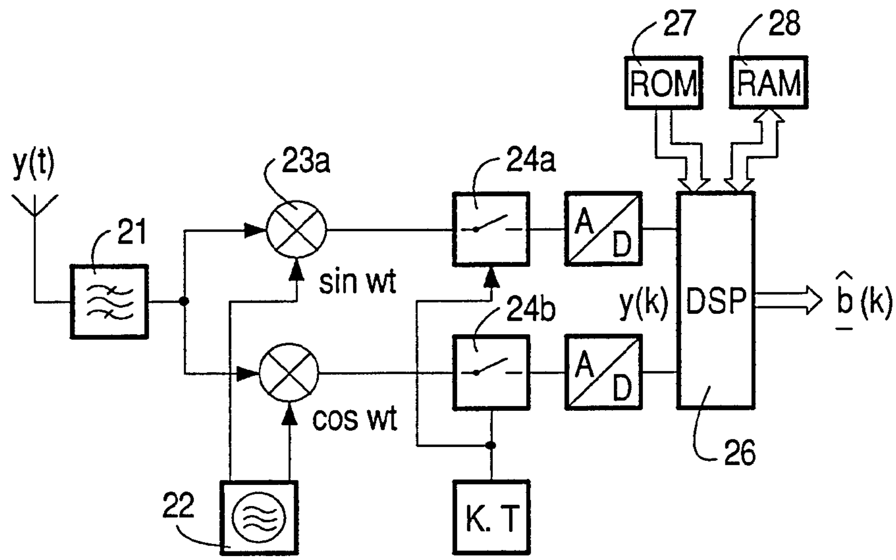
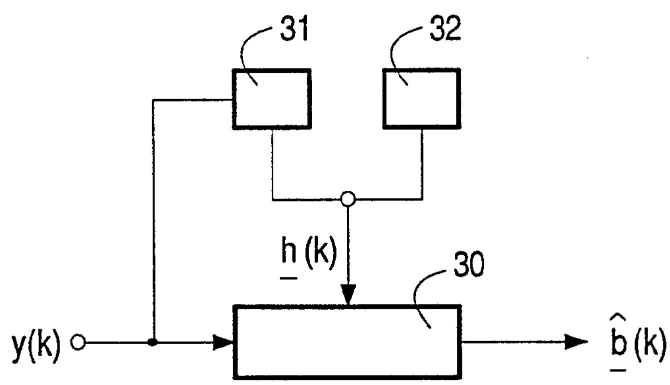


Fig. 1



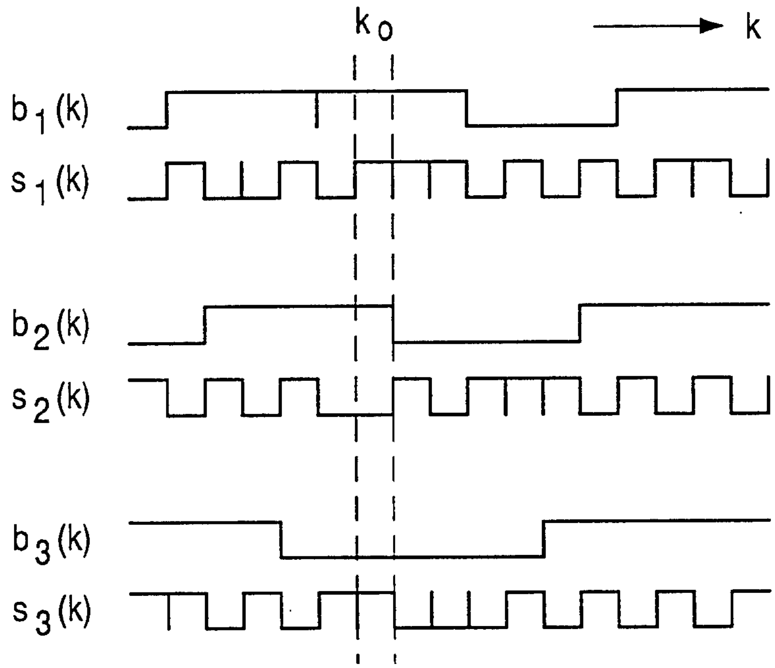
27
28
ROM
RAM
y(t)
23a
24a
21
A
D
sin wt
DSP
y(k)
24b
$\hat{b}$(k)
A
D
cos wt
26
22
K. T

Fig. 2



31
32
h(k)
30
y(k)
$\hat{b}$(k)

Fig. 3



$\underline{b}(\kappa_0) = [+1\ +1\ -1]^T$ $\underline{h}(\kappa_0) = [+1\ -1\ +1]^T$

Fig. 4

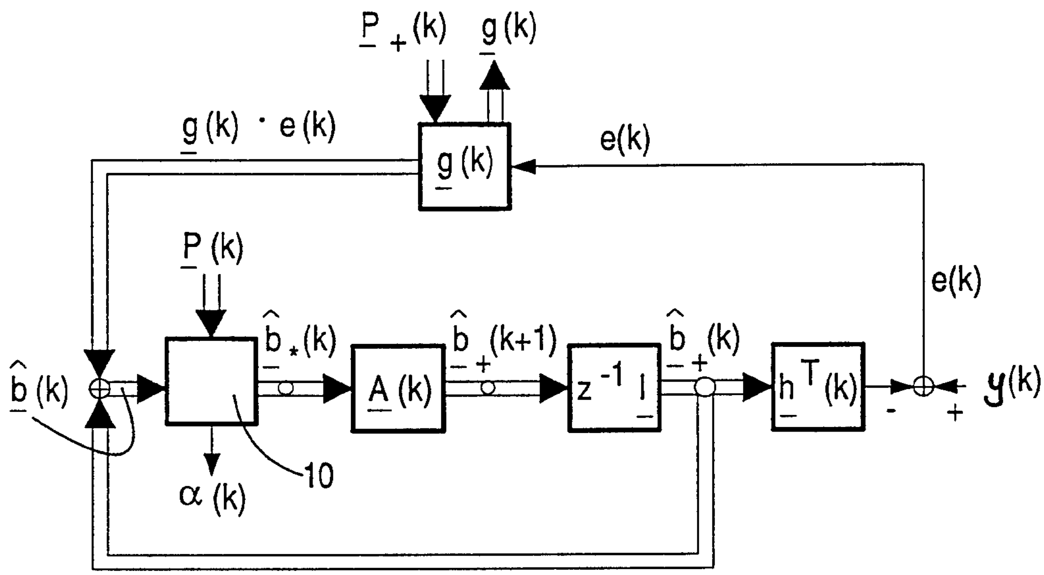

P + (k)
g (k)
g (k) · e (k)
g (k)
e(k)
e(k)
P (k)
b̂ * (k)
b̂ + (k+1)
b̂ + (k)
b̂ (k)
A (k)
z -1 I
h T (k)
y(k)
-
+
α (k)
10

Fig. 5

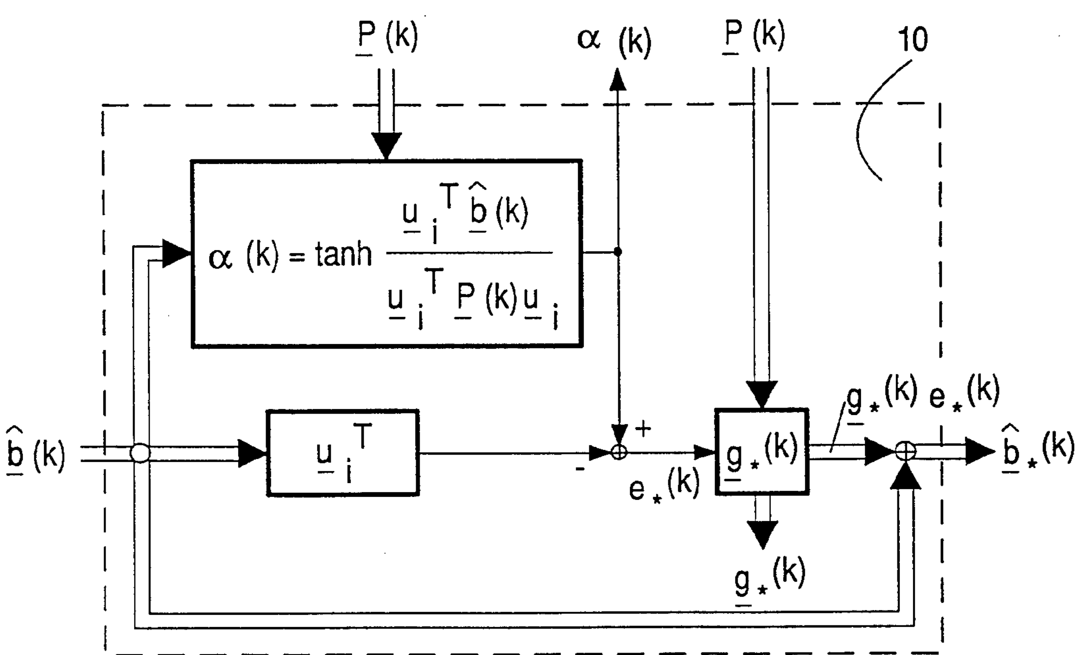

P (k)
α (k)
P (k)
10
α (k) = tanh (u i T b̂ (k)) / (u i T P (k) u i)
b̂ (k)
u i T
+
-
e * (k)
g * (k)
g * (k) e * (k)
b̂ * (k)
g * (k)

Fig. 6

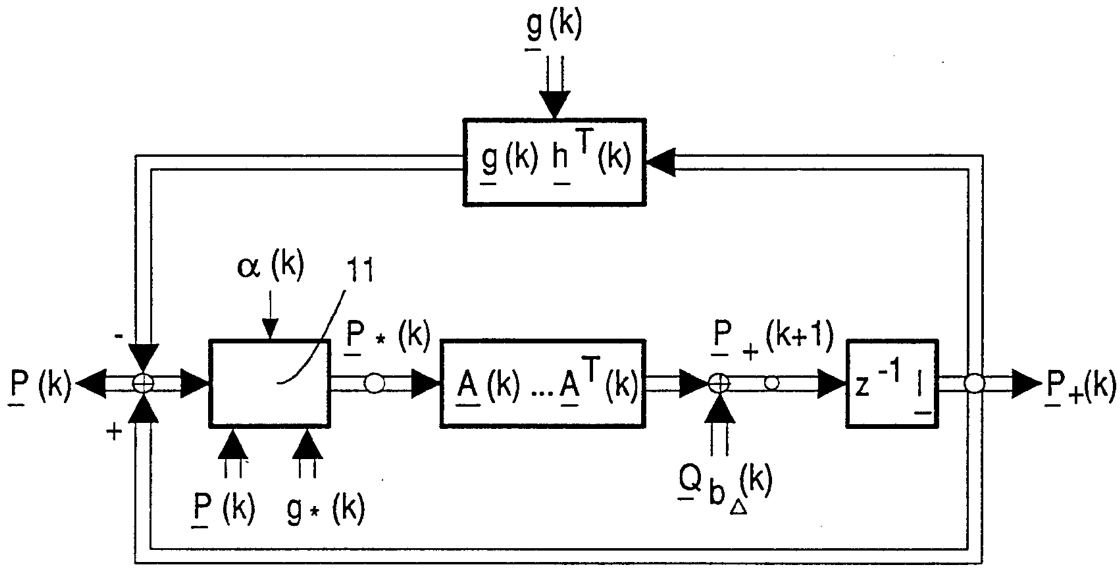

g (k)
g (k) h T (k)
α (k)
11
P * (k)
P + (k+1)
P (k)
A (k) ... A T (k)
z -1 I
P+(k)
-
+
P (k)
g * (k)
Q b△(k)

Fig. 7

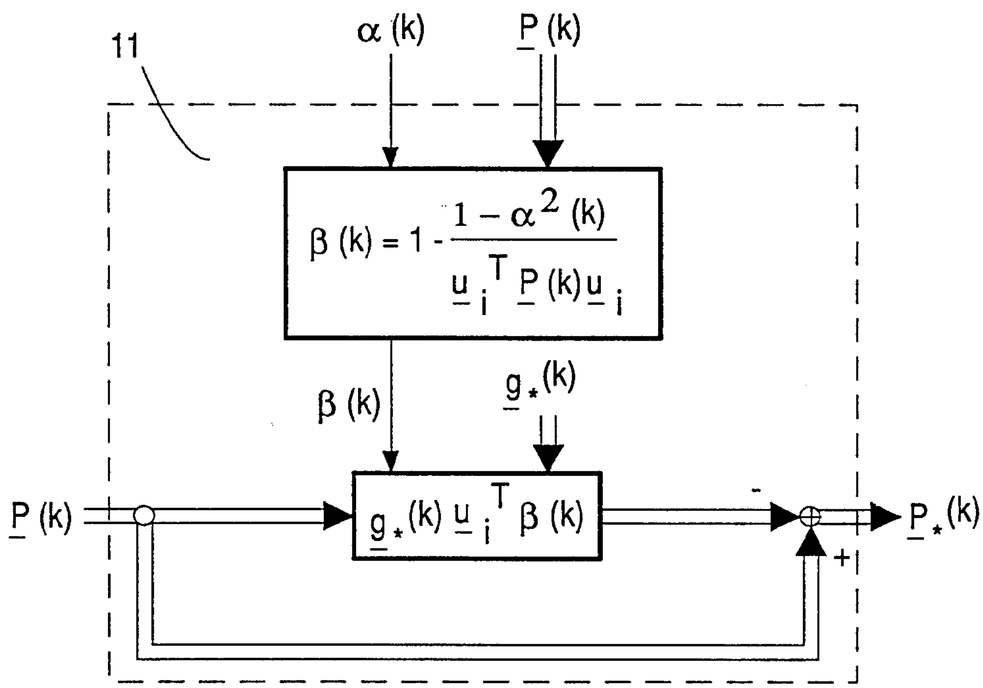

11
α (k)
P (k)
β (k) = 1 - (1 − α² (k)) / (u_i^T P (k) u_i)
β (k)
g * (k)
P (k)
g * (k) u i T β (k)
-
+
P * (k)

Fig. 8

INTERNATIONAL SEARCH REPORT

International application No.
PCT/IB 94/00116

A. CLASSIFICATION OF SUBJECT MATTER

IPC5: H04B 7/216
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

IPC5: H04B, H04J, H04Q

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

CLAIMS, WPI, EPODOC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US, A, 5099493 (A.E. ZEGER ET AL), 24 March 1992 (24.03.92), column 2, line 32 - column 3, line 30	1-10
	--	
A	US, A, 5166953 (J.E. HERSHEY ET AL), 24 November 1992 (24.11.92), column 1, line 58 - column 2, line 39, abstract	1-10
	--	
P,A	WO, A1, 9400926 (MOTOROLA INC.), 6 January 1994 (06.01.94), abstract	1-10
	--	

☐ Further documents are listed in the continuation of Box C. ☒ See patent family annex.

* Special categories of cited documents:
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"O" document referring to an oral disclosure, use, exhibition or other means
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"Y" document of particular relevance: the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
1 Sept 1994	08 -09- 1994

Name and mailing address of the ISA/	Authorized officer
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INTERNATIONAL SEARCH REPORT
Information on patent family members

02/07/94

International application No.
PCT/IB 94/00116

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
US-A- 5099493	24/03/92	NONE	
US-A- 5166953	24/11/92	NONE	
WO-A1- 9400926	06/01/94	NONE	